

Design of Transistor Regulated Power Supplies*

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Summary—A new form of transistor series regulated power supply is presented which permits unusually good performance characteristics to be realized with simple and economical circuitry. Expressions are given for the open-circuit output voltage and the output resistance in terms of the supply voltages and the circuit parameters. Application of the new circuit to two practical laboratory regulated power supplies is described, the specifications of which are: 0.5 ampere at 18 to 22 volts, and 1 a at 5 to 25 v. The output resistance is of the order of 0.01 ohm for both supplies; for sudden change of load current between minimum and maximum the transient in output voltage is 80 mv or less, and decays in 40 μ sec or less. At full load, ripple in the output voltage is less than 5 mv and line voltage variations of ± 10 per cent produce output variations of less than 5 mv. Each supply uses only one line transformer.

I. INTRODUCTION

THE DESIGN of vacuum tube regulated power supplies is well-established. Such supplies are suited for powering equipment in which relatively low currents at high voltages are required. However, for high-current low-voltage applications, vacuum tube supplies become bulky and inefficient, and the design is conveniently implemented with transistors instead of with tubes. Although various types of transistor regulated power supplies have been described¹⁻⁵, most of these possess comparatively low current ratings and poor transient response unless rather complex circuitry is used.

This paper presents a new form of transistor series regulated power supply which permits unusually good performance characteristics to be realized with simple and economical circuitry. Expressions are given for the regulated output voltage and the output resistance in terms of the supply voltages and circuit parameters.

Practical circuits for two laboratory regulated power supplies are presented. The input in each case is 110 v 60 cps and the other specifications are as follows:

Supply No. 1: 0.5 a at 18 to 22 v, output resistance 0.03 ohm.

Supply No. 2: 1 a at 5 to 25 v, output resistance <0.01 ohm.

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² F. H. Chase, "Power regulation by semiconductors," *Elec. Eng.*, vol. 75, pp. 818-822; September, 1956.

³ J. W. Keller, "Regulated transistor power supply design," *Electronics*, vol. 29, pp. 168-171; November, 1956.

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The transient response of each supply is excellent: sudden increase of load current from zero to maximum produces a transient in output voltage of less than 80 mv with a recovery time of about 10 μ sec; sudden decrease of load current from maximum to zero produces a transient in output voltage of less than 40 mv with a recovery time of about 40 μ sec. At full load, ripple in the output voltage is less than 5 mv and line voltage variations of ± 10 per cent produce output voltage variations of less than 5 mv. Each supply uses only one line transformer.

II. QUALITATIVE DEVELOPMENT OF THE NEW SERIES REGULATOR CIRCUIT

A simple form of standard series regulated power supply using *p-n-p* transistors is shown in Fig. 1. The voltage E is the regulated output and E_p is the unregulated "raw" input voltage with internal resistance R_p . It is required that the voltage E should be as constant as possible with respect to load current I for both steady-state and transient conditions, and should be as independent as possible of input voltage and circuit parameter variations.

If in Fig. 1, the base current of the transistor T_4 is neglected compared to the current through the potential divider, and if the base current of transistor T_1 is neglected compared to the collector current of T_4 , an equation for the output voltage may be simply derived as follows. In the steady state the output voltage E is given by

$$E = n(e + v) \quad (1)$$

where

e = the reference voltage developed across the zener diode D_1 ,

v = the emitter-to-base dc voltage of T_4 (approximately a tenth of a volt), and

n = ratio of the potential divider as defined in Fig. 1.

At any given load current I , there will be a voltage drop between the emitter and base of T_1 which is approximately $R_e I$, where R_e is the effective emitter-to-base resistance of T_1 . The voltage v at the input of T_4 adjusts itself so that the collector current of T_1 produces a voltage drop V across resistance R so that

$$V = E_p - R_p I - E - R_e I \quad (2)$$

Combining (1) and (2) gives

$$E = \left(ne + \frac{E_p}{G} \right) - \left(\frac{R_p + R_e}{G} \right) I \quad (3)$$

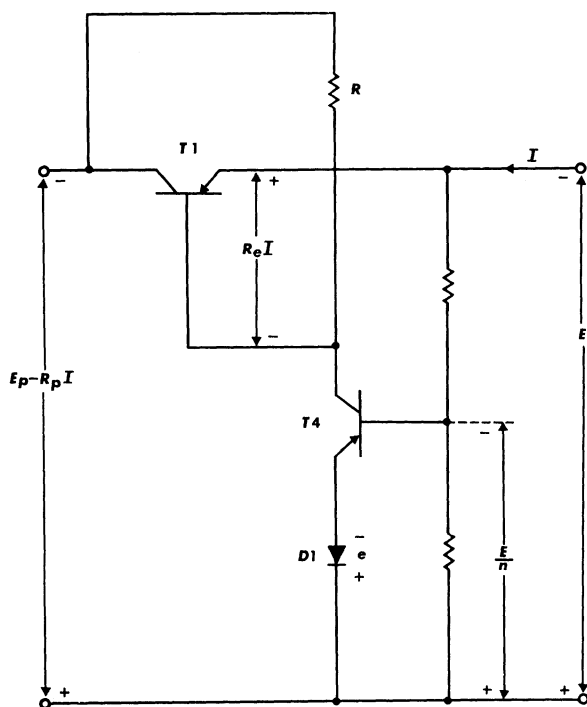


Fig. 1—Simple series regulator circuit, in which E is the regulated output voltage, and the supply voltage is E_p with internal resistance R_p . The output resistance is approximately $(R_e + R_p)/G$.

where $G = V/nv$, and is the dc gain of $T4$ and the potential divider. It is assumed that $G \gg 1$ in the derivation of (3).

In an ideal regulator $E = ne$, and thus (3) indicates the extent to which the circuit of Fig. 1 fails to be ideal. The undesired terms in (3) show that a practical regulator exhibits a finite output resistance and that the open-circuit output voltage is partially dependent on the raw supply voltage. In a typical case, suppose that the power supply is to provide zero to 1 a at 20 v and let $e = 5$ v, $R_e = 1$ ohm, $R_p = 9$ ohm, and $G = 10$. At maximum load of 1 a, 9 v will be dropped in the internal resistance R_p of the raw supply, so in order to maintain at least 1 v reverse bias between the base and collector of $T1$, the supply voltage E_p must be at least 30 v. To provide $E = 20$ v with $e = 5$ v requires $n = 4$ if the regulator approaches ideal performance. Substitution of the given figures into the undesired terms of (2) shows that $E_p/G = 3$ v, $(R_p + R_e)/G = 1$ v at 1 a load current. Thus if there is any variation in E_p (such as change in supply voltage or ripple) the variation in the output voltage E is 3/20 of the variation in E_p . The factor $(R_p + R_e)/G = 1$ ohm is the output resistance R_0 of the regulated supply, which is an improvement by about a factor of 10 over the output resistance R_p of the raw supply.

However, an output resistance of 1 ohm is still too high for many applications. A circuit modification which will further reduce the output resistance presents itself after a qualitative examination of how the circuit of Fig. 1 operates when the load current I is changed.

In the circuit of Fig. 1, an increase in load current I tends to cause a drop in output voltage E not only be-

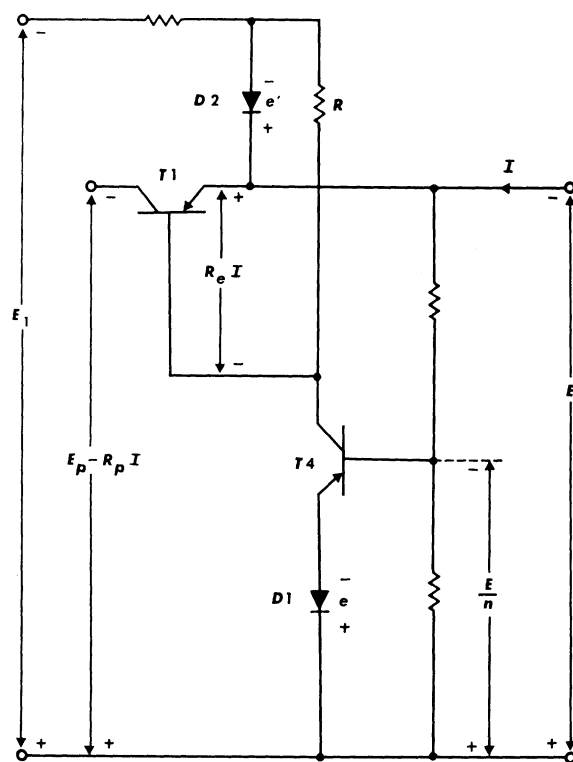


Fig. 2—Principle of the new series regulator circuit, in which the output resistance is reduced approximately to R_e/G .

cause of the larger drop $R_e I$ in $T1$, but also because of the larger drop $R_p I$ in the internal resistance of the raw supply E_p which is connected to the output via the resistor R and transistor $T1$. Thus the amplifier $T4$ is required to compensate for both these drops, and hence the factor $(R_p + R_e)$ occurs in (3). Obviously, if some means could be found to maintain the voltage at the top end of R constant with respect to changes in load, the output resistance would be equal to R_e/G instead of to $(R_p + R_e)/G$, or 0.1 ohm instead of 1 ohm for the above numerical values. This may be accomplished by the circuit of Fig. 2, in which the top end of R is connected to a voltage more negative than the output voltage by a fixed amount. The presence of the Zener diode $D2$ insures that the voltage at the top end of R is always essentially constant for all load currents, and thus the above conditions for lower output resistance are realized.

An expression for the output voltage E in the circuit of Fig. 2 is easily obtained by substituting $E + e'$ for $E_p - R_p I$ in (3), which leads to

$$E = ne + \frac{e'}{G} - \frac{R_e}{G} I. \quad (4)$$

It is seen that the output resistance R_0 is indeed R_e/G , or 0.1 ohm for the figures given previously. Further, the output voltage is independent of E_p and the auxiliary supply E_1 , and thus E will be independent of supply voltage variations or ripple.

It is apparent that the circuit of Fig. 2 provides a marked improvement over that of Fig. 1, and yet re-

quires only two more components and an extra (low-current) supply voltage. The simple semiquantitative discussion above is sufficient to explain the principle of the modified regulated supply circuit of Fig. 2, but in practical circuits several additional factors must be taken into account. The most important of these are that the zener reference diodes $D1$ and $D2$ possess finite dynamic resistance (thus the voltages e and e' are not quite constant for different currents), and that for maximum load currents in excess of a few milliamperes the base current of $T1$ will certainly not be negligible compared to the collector current of $T4$. All these factors affect the gain G of $T4$, and also cause the output voltage not to be completely independent of the supply voltage. A more detailed analysis of this new circuit follows in the next section.

III. ANALYSIS OF THE REGULATOR CIRCUIT

A practical embodiment of the principle described in the previous section is shown in Fig. 3. The most important change is the addition of two more transistors $T2$ and $T3$, connected as emitter followers, to provide current amplification so that load currents of several amperes may be drawn, and yet to keep the base current of $T3$ small enough to be comparable with or smaller than the collector current of $T4$. A base current in $T3$, which is not negligible compared to the collector current of $T4$, means that the input resistance of $T3$ shunts appreciably the resistance R . The three transistors $T1$, $T2$, and $T3$ will usually be adequate for load currents up to 1 a. An ammeter will normally be required to monitor the load current, and must be inserted inside the feedback loop; otherwise, the output resistance of the supply will be increased by the resistance of the meter.

If the dc current through the transistor $T4$ is insufficient to maintain the reference diode $D1$ in the breakdown region, the auxiliary supply E_2 may be necessary to provide extra current. It is desirable that E_2 be obtained from the same line transformer that supplies the external load current, in which case E_2 will contain a ripple component and will vary with load current I owing to the imperfect regulation of the raw supply. This effect may be accounted for by supposing the supply E_2 to contain a "transfer internal resistance" R_2 , which may be of the order of 10 ohms. The other auxiliary supply E_1 will possess in a similar way a transfer internal resistance R_1 .

The resistance R_f is very small, of the order of a fraction of an ohm, and is inserted to provide positive feedback proportional to the load current so that the output resistance of the regulated supply can be made zero or slightly negative. The feedback resistance may not always be necessary, but will be included in the analysis so that its effect can be evaluated.

Analysis of the circuit of Fig. 3 requires finding the equation corresponding to (4) for the regulated output voltage, and is conveniently broken down into two

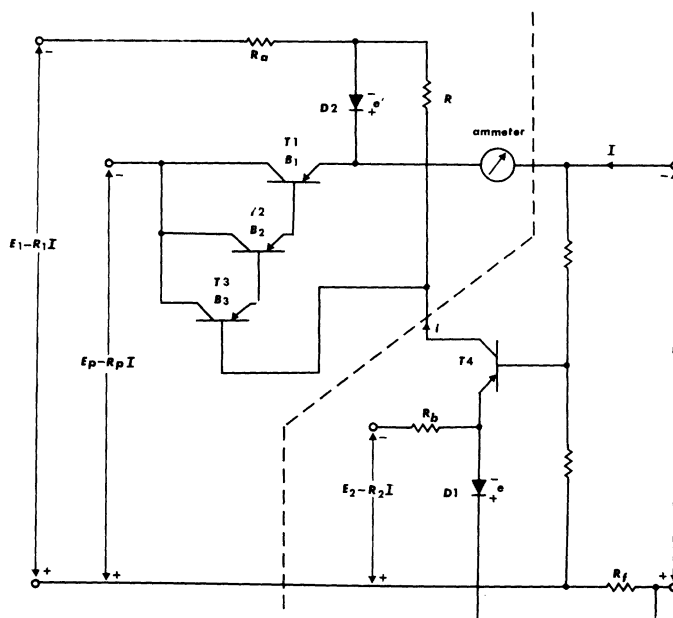


Fig. 3—Practical embodiment of the circuit principle of Fig. 2.

parts. First, an expression for the collector current i of transistor $T2$ is found in terms of the output voltage E by consideration of the voltage amplifier part of the circuit to the right of the dashed line in Fig. 3. Second, an expression for E in terms of i is obtained by consideration of the current amplifier part of the circuit to the left of the dashed line in Fig. 3. Elimination of i from the two resulting equations leads to the desired expression for the output voltage E :

$$E = \left[ne + \frac{r}{R_b} nE_2 + \frac{r'}{R_a} \frac{1}{G} E_1 + \frac{1}{G} e' \right] - I \left[\frac{(R_{e1} + R_{e2} + R_{e3}) + R/B_1B_2B_3}{G} + \frac{R_1 + R_m}{(R_a/r')G} + \frac{r}{R_b} nR_2 - \frac{1+n}{n} R_f \right] \quad (5)$$

where

e, r = the reference voltage, dynamic resistance of zener diode $D1$,

e', r' = the corresponding quantities of zener diode $D2$,

R_m = the resistance of the ammeter,

B_1, B_2, B_3 = the common-emitter dc current gains of $T1, T2, T3$,

R_{e1}, R_{e2}, R_{e3} = effective emitter-base resistances of $T1, T2, T3$ (defined in each case as the ratio: change in emitter-base voltage/change in load current I),

n = the ratio of the potential divider, defined as in Fig. 1,

G = is the dc gain of the potential divider and the amplifier $T4$,

and the other quantities are as shown in Fig. 3. The gain G corresponds to the quantity V/nv previously

mentioned, and is given by

$$G = R/n[r_e + r + (r_p + r_b)/B_4] \quad (6)$$

where r_p is the resistance of the two potential divider resistors in parallel, and r_e , r_b , and B_4 are the emitter resistance, base resistance, and common-emitter dc current gain of T_4 .

In the derivation of (5), the following good approximations are made:

$$B \gg 1 \text{ for all transistors} \quad (7)$$

$$R_b \gg r \quad (8)$$

$$R_a \gg r' \quad (9)$$

$$R \gg r' \quad (10)$$

$$(R_a/r')G \gg 1. \quad (11)$$

It is further assumed that the load current I is considerably larger than the current drain through the potential divider, and is also considerably larger than the current through the diode D_2 .

Eq. (5) may be written

$$E = E_0 - R_0 I \quad (12)$$

where E_0 is the open-circuit output voltage and R_0 is the effective output resistance of the regulated supply. The criteria of a satisfactory design are that E_0 should be as independent as possible of the raw supply voltages E_p , E_1 , and E_2 , and the output resistance should be as small as possible (less than 1 ohm). Eq. (5) shows the relative effects of the raw supply voltages and the various internal resistances which cause the output voltage to deviate from the ideal value $E = ne$. A discussion of the implications of (5) is desirable, and is given in the next section.

IV. DISCUSSION OF THE ANALYTICAL RESULTS

Consider first the effects of the raw supply voltages on the open-circuit output voltage E_0 , where, from (5) and (12),

$$E_0 = ne + \frac{r}{R_b} nE_2 + \frac{r'}{R_a} \frac{1}{G} E_1 + \frac{1}{G} e'. \quad (13)$$

In an ideal regulator $E_0 = ne$, and thus the other terms in (13) are undesirable and should be minimized in a good design. In order to obtain a high value of the ratio R_a/r' , the supply E_1 should be at least twice the output voltage E . Let typical figures again be taken of $e = 5$ v, $n = 4$; then E_0 will be equal to 20 v plus correction terms. The supply E_1 should be at least 40 v, and for $G = 10$ and $R_a/r' = 100$ the term in E_1 is equal to 0.04 v, or 0.2 per cent of the output voltage. Thus E_0 is almost completely independent of E_1 , and practically none of any ripple on E_1 will get through to the output.

In some cases the supply E_2 will not be required at all, since sufficient dc current to break down the reference diode D_1 may be available through transistor T_4 . In this case, of course, the terms in E_2 will not appear in the

equation for E_0 . However, if extra current is required for the diode D_1 , it may be obtained from the supply E_p in which case $E_2 = E_p$. For $E_p = 30$ v, the ratio R_b/r may be about 100, and so the term in E_2 in (13) will be equal to 1.2 v. This represents 6 per cent of the output voltage E_0 , a figure which is undesirably high since variations in line voltage and ripple will thus be transmitted to the regulated output with appreciable magnitude. A way to overcome this problem is to make E_2 equal not to E_p but to E ; that is, to derive the extra current for D_2 from the regulated output instead of from the raw input. Putting $E_2 = E$ in (5) leads to

$$E_0 = \frac{1}{1 - nr/R_b} \left(ne + \frac{r'}{R_a} \frac{1}{G} E_1 + \frac{1}{G} e' \right) \quad (14)$$

and the result is essentially the same as when $E_2 = 0$ except that the no-load output voltage is increased by the factor $1/(1 - nr/R_b)$ which, however, is close to unity.

The remaining term to consider in (5) is that in e' . If $e' = 5$ v, this term amounts to 2 per cent of the output voltage and is not detrimental since e' is itself a reference voltage.

To summarize the above discussion of the open-circuit output voltage, the typical numerical values obtained above are shown below in a form in which the terms correspond to those in (14), where $E_2 = E$:

$$E_0 = \frac{1}{1 - 0.04} (20 + 0.04 + 0.5) \quad (15)$$

$$E_0 = 21.36 \text{ v.} \quad (16)$$

Consider next the effects of the various internal resistances on the output resistance R_0 . From (5) and (12),

$$R_0 = \frac{(R_{e1} + R_{e2} + R_{e3}) + R/B_1 B_2 B_3}{G} + \frac{R_1 + R_m}{(R_a/r')G} + \frac{r}{R_b} nR_2 - \frac{1+n}{n} R_f. \quad (17)$$

In an ideal regulator R_0 should be as small as possible, and hence all the terms in (17) should be minimized in a good design. Each term which contributes to the output resistance R_0 can be identified with an independent process occurring in the circuit of Fig. 3.

The quantity $(R_{e1} + R_{e2} + R_{e3})$ is the total effective internal resistance of the current-amplifying transistors T_1 , T_2 , and T_3 and corresponds to the internal resistance R_e of the single transistor shown in the circuit of Fig. 2. As predicted by the simple analysis given in connection with Fig. 2, the internal resistance of the current-amplifying transistor contributes a term $(R_{e1} + R_{e2} + R_{e3})/G$ to the output resistance of the supply. However, in the simple analysis it was assumed that the base current of the current-amplifying transistor was negligible compared to the current through R , that is, the resistance looking into the base of the current-amplifying transistor was much larger than the resistance R . In the

more accurate analysis given above this assumption was not made, and the presence of this shunting effect is represented by the term in R in (17).

Some typical figures again will help to give a feeling for the relative magnitudes of the terms in (17) for the output resistance. Suppose that $G=10$, $R=6000$ ohms, $B_1, B_2, B_3=30, 40, 50$, and $R_{e1}, R_{e2}, R_{e3}=0.70, 0.15, 0.15$ ohms respectively. The total effective internal resistance R_e of transistors $T1, T2$, and $T3$ is then 1 ohm, and the additional resistance contributed by the term in R is 0.1 ohm. Hence, the component of output resistance arising from these two sources is $(1+0.1)/10=0.11$ ohm, which may be compared with the result of 0.10 ohm which would be calculated by neglecting the shunting effect.

From (17) it is seen that an additional component of output resistance arises from the presence of the ammeter resistance R_m , and of the auxiliary supply transfer internal resistance R_1 . Typical values are $R_m=1$ ohm, $R_1=10$ ohm. However, it is seen that the resulting component of output resistance is the sum of the two resistances divided not by G alone, but by $(R_a/r')G$. For the above typical figures, this component of output resistance is 0.011 ohm, and is only 10 per cent of the component due to R_e and R .

A third component of output resistance is introduced by the transfer internal resistance R_2 of the auxiliary supply E_2 . As described in connection with the open-circuit output voltage E_0 , the supply E_2 will not be required in some cases, and so $R_2=0$. However, if E_2 is taken from the main supply voltage E_p , $R_2=R_p$ and may be about 10 ohm. For $R_b/r=100$, $n=4$ as before, the component of output resistance due to R_2 will be 0.4 ohm and is undesirably large. This problem may be resolved by again deriving the extra current supply for diode $D1$ from the output voltage E instead of from E_p . In this case $R_2=0$, and putting $E_2=E$ in (5) gives

$$R_0 = \frac{1}{1 - nr/R_b} \left[\frac{(R_{e1} + R_{e2} + R_{e3}) + R/B_1B_2B_3}{G} + \frac{R_1 + R_m}{(R_a/r')G} - \frac{1+n}{n} R_f \right] \quad (18)$$

and the result is essentially the same as when E_2 and R_2 are zero, except that the output resistance is increased by the factor $1/(1-nr/R_b)$ which, however, is close to unity.

The remaining term to consider in (17) is that due to the feedback resistance R_f . As described in Section III, R_f introduces positive current feedback whose magnitude can be adjusted to make R_0 zero or even negative. From (18) the value of R_f which would be required to make $R_0=0$ for the above typical figures, and where $E_2=E_p$ and $R_2=R_p$, is given by

$$0 = 0.11 + 0.011 + 0.4 - (5/4)R_f \quad (19)$$

or

$$R_f = 0.417 \text{ ohm.} \quad (20)$$

A resistance of this small magnitude is furnished by a few inches of wire, and, in practice, the best way to obtain the desired value of R_f is to adjust the length of a piece of wire in the appropriate place in the completed regulated supply circuit until the output resistance is zero. This method of reducing the output resistance is not particularly satisfactory, however, since the adjustment is awkward to make and in any case will only be correct for a particular value of n . Thus, if a variable output voltage E is required (which is obtained by varying the potential divider ratio n), the output resistance R_0 may vary considerably over the range of output voltage. Therefore, it is preferable to design for a low inherent output resistance by making G as large as possible.

To summarize the above discussion of the output resistance, the typical numerical values obtained above are shown below in a form in which the terms correspond to those of (18), in which $E_2=E_1$, $R_2=0$, and where $R_f=0$:

$$R_0 = \frac{1}{1 - 0.04} (0.11 + 0.011) \quad (21)$$

$$= 0.126 \text{ ohm.} \quad (22)$$

V. A FURTHER CIRCUIT MODIFICATION

There is yet another beneficial modification which may be made in the regulated power supply circuit, and which is suggested by (15) for the output voltage.

It may be noted that in (15) there is a component of output resistance due to the total effective internal resistance $R_e=(R_{e1}+R_{e2}+R_{e3})$ of the current-amplifying transistors $T1, T2$, and $T3$, and another component due to the resistance R_m of the ammeter. The former component is equal to R_e divided by G , but the latter component is R_m divided by $(R_a/r')G$ where R_a/r' is at least 100. Examination of Fig. 3 suggests that if the diode $D2$ were connected not to the emitter of transistor $T1$ but to the emitter of $T2$, then the two above components of output resistance might be equal to $(R_{e2}+R_{e3})/G$ and $(R_m+R_{e1})/(R_a/r')G$, respectively, and the sum of the components and hence the output resistance R_0 would be reduced.

If the above modification is introduced, analysis corresponding to that for Fig. 3 leads to

$$E = \left[ne + \frac{r}{R_b} nE_2 + \frac{r'}{R_a} \frac{1}{G} E_1 + \frac{1}{G} e' \right] - I \left[\frac{(R_{e2} + R_{e3}) + R/B_1B_2B_3}{G} + \frac{R_1 + R_m + R_{e1}}{(R_a/r')G} + \frac{r}{R_b} nR_2 - \frac{1+n}{n} R_f \right]. \quad (23)$$

The result of (23) shows that the expectation described above is justified: the open-circuit output voltage E_0 is the same as before, but in the output resistance R_0 , the resistance R_{e1} is now divided by $(R_a/r')G$ instead of by

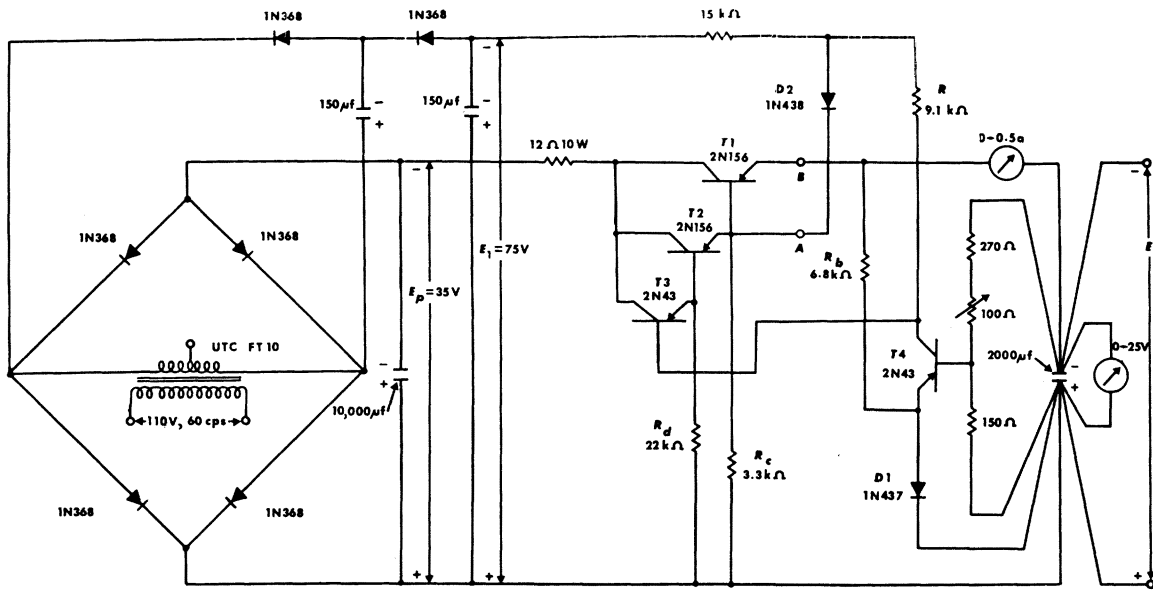


Fig. 4—Circuit of regulated supply to provide 0.5 a at 18 to 22 v. The output resistance is 0.007 ohm.

G alone. The same typical figures assumed before will show the improvement introduced by the circuit modification. If $G = 10$, $R = 6000$ ohm, $B_1, B_2, B_3 = 30, 40, 50$, and $R_{e1}, R_{e2}, R_{e3} = 0.70, 0.15, 0.15$ ohm, respectively, then

$$\frac{(R_{e2} + R_{e3}) + R/B_1B_2B_3}{G} = \frac{0.3 + 0.1}{10} \quad (24)$$

$$= 0.04 \text{ ohm} \quad (25)$$

and

$$\frac{R_1 + R_m + R_{e1}}{(R_a/r')G} = \frac{10 + 1 + 0.7}{100 \times 10} \quad (26)$$

$$= 0.0117 \text{ ohm}. \quad (27)$$

and if E_2, R_2 , and R_f are all zero the output resistance R_0 is

$$R_0 = 0.04 + 0.0117 \quad (28)$$

$$\approx 0.052 \text{ ohm} \quad (29)$$

which may be compared with $R_0 = 0.121$ ohm obtained with the circuit of Fig. 3.

Obviously, whether or not the circuit modification introduces a significant reduction in R_0 depends on the actual numerical values in the circuit. For example, the improvement will not be significant if there is appreciable shunting of R , that is, if $R/B_1B_2B_3$ is comparable to $(R_{e1} + R_{e2} + R_{e3})$, or if R_{e1} is comparable with or smaller than $(R_{e2} + R_{e3})$. This latter condition may occur if two transistors are used in parallel in place of the single transistor $T1$ in order to obtain larger maximum load current I .

The discussion of the preceding two sections has presented the analysis and interpretation of the results to be expected from a new form of series-regulated power supply circuit. Although the treatment has been in

terms of a fixed output voltage E , there is nothing in the principle which prevents extension to variable-voltage supplies. However, certain practical problems do arise if the output voltage is required to be varied by more than about ± 10 per cent. In the following sections two practical power supplies embodying the above principle are described, one having an output voltage variable by ± 10 per cent, and one having an output voltage variable over a wider range.

VI. PRACTICAL SUPPLY NO. 1: 0.5 A AT 18 TO 22 V

The schematic diagram of a regulated power supply to provide up to 0.5 a at 18 to 22 v is shown in Fig. 4. The input power source is 110 v 60 cps.

The raw supply voltage E_p is obtained from a bridge rectifier circuit, and the auxiliary supply voltage E_1 is obtained from a voltage-doubler from the same line transformer. The auxiliary supply E_2 is not required; instead, extra bias current for the zener diode $D2$ is obtained from the output voltage through resistor R_b , as described in Section IV.

The two resistors R_e and R_d are introduced so that the transistors $T2$ and $T3$, respectively, will not be cut off when the load current is small. The transient response is also improved because the current-amplifying transistors are working in a linear region over a greater part of their range. The presence of the resistors R_e and R_d modifies the details of the analysis presented in the previous sections, but the principle remains the same.

The 2000- μ f capacitor connected across the regulated output is added to improve the transient response. The wiring layout indicated in Fig. 4 is particularly important: all leads should be returned directly to the capacitor terminals as shown. The reason is that the load current must not be allowed to develop undesired voltages which find their way to the amplifier input. As described in Section IV, the presence of a feedback re-

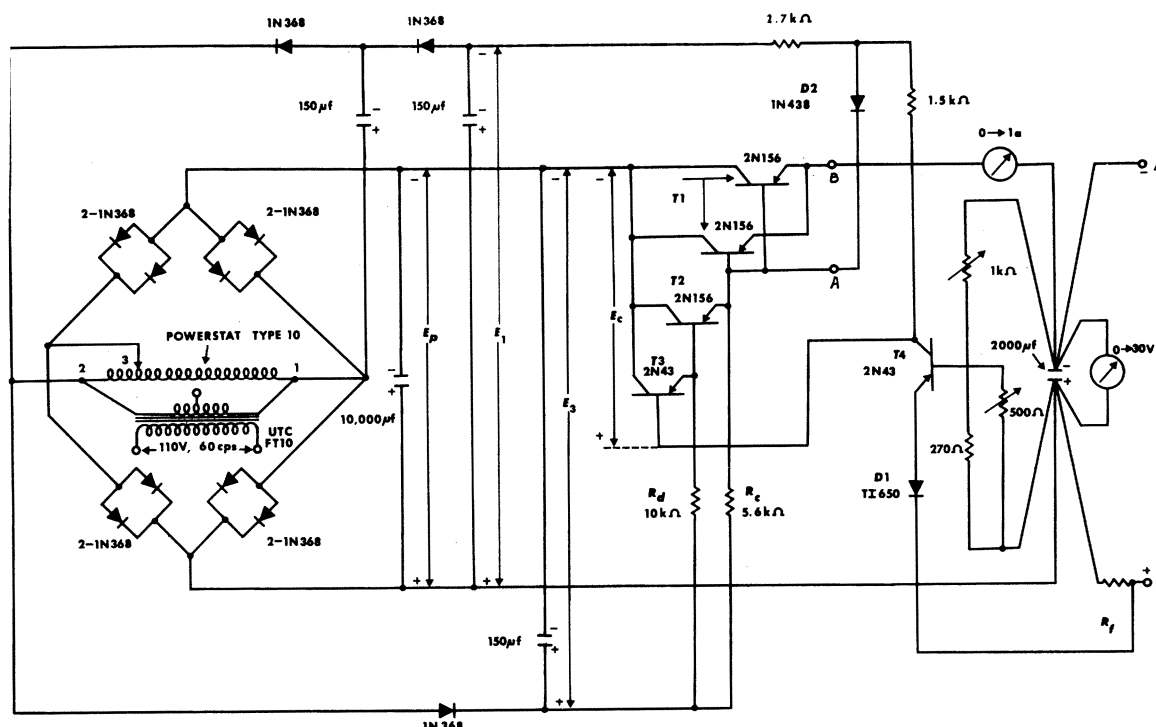


Fig. 5—Circuit of regulated supply to provide 1 a at 5 to 25 v. The output resistance can be made zero at the maximum output voltage.

sistance R_f of only a fraction of an ohm in the right place can cause an appreciable change in output resistance.

The presence of the resistors R_c and R_d , the difficulty in measuring the dynamic resistances of the reference diodes, and the difficulty in predicting the gain of the voltage amplifier make it rather impracticable to calculate the value of the output resistance to be expected. Experimental measurements indicate the following performance characteristics: for a suddenly applied load of 0.5 a, the output voltage drops 40 mv and recovers in about 10 μ sec; for a suddenly removed load of 0.5 a the output voltage rises 20 mv and recovers in about 40 μ sec. Ripple and noise in the output are less than 5 mv at full load, and the output voltage changes by less than 5 mv for line voltage changes of ± 10 per cent.

With the diode $D2$ connected to point B instead of to point A, the steady-state output resistance is 0.03 ohm. With the diode connected to point A the output resistance is 0.007 ohm, showing that significant improvement is obtained by the circuit modification described in Section V.

VII. PRACTICAL SUPPLY NO. 2: 1 A AT 5 TO 25 V

The design of a regulated power supply to provide an output voltage variable over a wide range poses several problems in addition to those already discussed for small-range supplies. The most important problem is that at low output voltage settings the power dissipation in the current-amplifying transistor becomes excessive. This can, of course, be alleviated by putting more transistors in parallel, but this is a wasteful and expensive procedure. A more attractive solution is to reduce

the raw dc supply voltage E_p by means of a Powerstat, at the same time that the output voltage E is reduced, and then the power dissipation in the series transistors remains small.

The schematic of a regulated power supply to provide up to 1 a at any voltage between 5 and 25 v is shown in Fig. 5. The input power source is 110 v 60 cps. The Powerstat is ganged to the 1-kilohm rheostat which sets the output voltage. The regulator part of the circuit is similar in principle to the corresponding part of the 20-v supply described in the previous section. However, the raw dc supplies in Fig. 5 are of special interest.

Three input voltages are provided: the heavy-current supply E_p , obtained from a bridge rectifier; the auxiliary supply E_1 to bias the zener diode $D2$, obtained from a voltage doubler; and another auxiliary supply E_3 to provide current through the resistors R_c and R_d , obtained from a half-wave rectifier. Fig. 6 shows the variation of the open-circuit characteristics of these three input voltages with Powerstat setting. The purpose of arranging the inputs in this way is to maintain all the diodes and transistors at the same operating point, and the same currents in resistors R_c and R_d , at all settings of the output voltage.

An additional problem encountered in a power supply to provide a wide range of output voltages is that extra current for biasing zener diode $D1$ cannot be obtained from the output voltage E as was done in the 20-v supply described previously. The reason is that since the output voltage may vary over a wide range, the extra current would vary likewise and in fact would fall to zero at the lowest output voltage. An alternative is

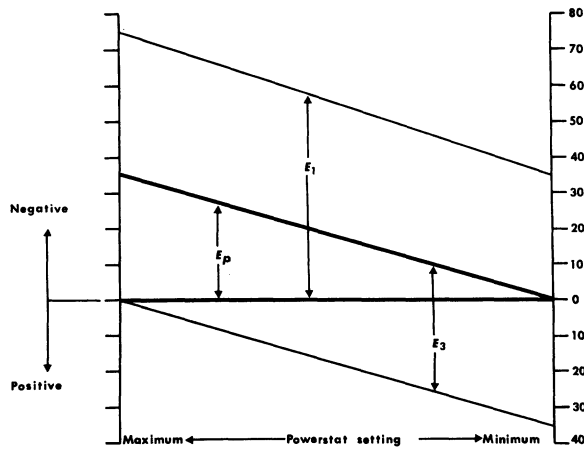


Fig. 6—Variation of the raw supply voltages, E_p , E_1 , and E_3 with Powerstat setting, in the circuit of Fig. 5.

to derive the extra current from another auxiliary supply E_2 (as described in Section III) which does not vary with the output voltage setting. However, this is unsatisfactory because of the unavoidably large transfer internal resistance R_2 of the supply E_2 , and because of the ripple introduced into the output. The only remaining possibility (short of using a second line transformer and well-filtered rectified output) is to allow sufficient current through the voltage-amplifying transistor to bias the diode $D1$ adequately without the necessity for extra current. In the circuit of Fig. 5, the resistance R is therefore made small in order to pass about 5 ma through the voltage-amplifying transistor. Consequently, the gain of this stage is considerably less than in the supply described previously.

In the absence of the resistance R_f in the circuit of Fig. 5, the output resistance is about 0.25 ohm at the maximum output voltage of 25 v, irrespective of whether the diode $D2$ is connected to point A or to point B. This rather large value of output resistance is principally due to the low gain of the voltage amplifier (because of the low value of R) and also to the high value of n (the potential divider ratio). However, the introduction of positive current feedback as described in Section III can be usefully employed to reduce the output resistance. The circuit must be carefully wired as indicated in Fig. 5, and the resistance R_f consists of about 30 inches of No. 22 tinned copper wire. The required length is best found by trial and error, and is chosen so that the output resistance is zero or slightly positive at maximum output voltage setting. At lower output voltages, the output resistance will then be almost zero or slightly negative. The transient response is comparable with that of the 20-v supply described in the previous section, the actual magnitudes involved being dependent on the output voltage setting.

VIII. CONCLUSION

A new form of transistor series regulated power supply circuit has been introduced and analyzed. Application of this new circuit to two practical regulated supplies has been described, one variable over a small range and one variable over a wide range of voltage. Each supply is simple and economical and requires only one line transformer. Independence of line voltage variations, output voltage ripple, output resistance, and transient response are unusually good.

CORRECTION

W. Guggenbuehl and M. J. O. Strutt, authors of "Theory and Experiments on Shot Noise in Semiconductor Junction Diodes and Transistors," which appeared on pages 839–854 of the June, 1957, issue of these PROCEEDINGS, have brought the following corrections to the attention of the editors.

P. 841, (5), should read: $\overline{i_{n \text{ tot}}^2}$.

P. 843, (12), should read: $\overline{i_{n \text{ th}}^2}$.

P. 843, three lines below (12), should read: $\overline{i_{nIe}^2} = 2eI\Delta f$.

P. 843, first column, second line from bottom, should read: $\overline{i_{nIc}^2}$.

P. 843, (13), Y Index should read: Y_L .

P. 845, first column, sixth line from top, should read:

$$|Y_{11}| \gg |Y_{12}|, |Y_{21}| \gg |Y_{22}|.$$

P. 846, first column, thirteenth line from top, should read: $\overline{i_{ne}^2}$.

P. 846, first column, fourteenth line from top, should read: $\overline{i_{nc}^2}$.

P. 846, second column, tenth line from bottom, should read: $\overline{u_{nh}^2}$.

P. 846, second column, ninth line from bottom, should read: $\overline{i_{nh}^2}$.

P. 847, first column, third line from bottom, should read:

$$(\overline{i_{n1}^* + i_{n2}^*})(i_{n1} + i_{ns} + i_{np}).$$

P. 854, Acknowledgment, second line from top, should read to the firms Brown, Boveri at Baden and Hasler at Berne, Switzerland.